A Method and Apparatus for Digital Near-End Echo/Near-End Crosstalk Cancellation with Adaptive Correlation

Background of the Invention

Field of the Invention

The present invention relates to apparatus and methods for communication of signals in a communication medium. More particularly, this invention relates to apparatus and methods to cancel echo interference and near-end crosstalk interference in a received signal from a transmitter attached to the communication system and from transmitters attached to other communication media in close proximity to the communication media.

Description of the Related Art

The gigabit Ethernet (1000 BASE-T) as defined by the IEEE standard 802.3ab is well known in the art. The structure capabilities and design consideration are described in:

"Gigabit Ethernet Over 4-Pair 100 OHM Category 5 Cabling," Gigabit Ethernet Alliance, Cupertino, CA, 1999,

"Gigabit Ethernet 1000 Base-T," 1000 BASE-T Tutorial Series, Interoperability Laboratory Gigabit Ethernet Consortium, University of New Hampshire, Durham, NH, 1998,

"Design Considerations for Gigabit Ethernet 1000 Base-T Twisted-pair Transceivers," Hatamian et al., Proceedings of the IEEE 1999 Custom Integrated Circuit Conference, IEEE, 1998, pp. 335-342.

Transmitting a gigabit data stream over four pair of category 5 unshielded twisted-pair cabling as described in the above-referenced papers has several design challenges. These challenges include signal attenuation, echo return loss, crosstalk characteristics of the cable, and electromagnetic emission and susceptibility.

Attenuation is the signal loss of the cabling from the transmitter to the receiver. Attenuation increases with frequency, which is due to such factors as skin effect. To minimize the effect of attenuation, the lowest possible frequency range that is consistent with the required data rate must be employed.

Echo is a by-product of the dual-duplex operation, where both the transmit and receive signal occupy the same wire pair. The residual transmit signal due to the trans-hybrid loss and the cabling return loss combine to produce an unwanted signal referred to here as echo.

Return loss is a measure of the amount of power reflected due to cabling impedance mismatches.

Crosstalk is an unwanted signal coupled between wire pairs that are in close proximity. Since 1000 BASE-T will use all four wire pairs, each pair is affected by crosstalk form the adjacent three pairs. Crosstalk is characterized in reference to the transmitter. Near-end crosstalk (NEXT) is crosstalk that appears at the output of a wire pair at the transmitter end of the cable and far-end crosstalk (FEXT) that appears at the output of a wire pair at the far end of the cable from the transmitter. Equal level far-end crosstalk (ELFEXT) is FEXT with the cable attenuation removed to provide equal level comparisons, i.e. crosstalk and receive signals voltages are compared at the end of the cabling opposite the transmitter. Crosstalk must be minimized to insure correct symbol recovery operations in the receiver.

A transmission system operating over unshielded cable must be capable of withstanding radiated energy from other sources, including AM, CB, short wave radio, and other external transmitters. The transmission system is required to have a tolerance to a 3 V/m continuous wave source above 27 MHz.

A further requirement is that the transmission system be immune to background and impulse noise. Impulse noise can be generated by power line transients, electrical fast transients, electrostatic discharge (ESD), and other sources.

Refer now to Fig. 1 for a discussion of near-end echo interference and near-end crosstalk. Fig. 1 shows a diagram of a gigabit Ethernet communications system. The gigabit Ethernet has two nodes that transmit and receive 1000 M bits per second (bps) full-duplex and bi-directionally. Each node consists of four transmitter/receivers (transceivers) **5a**, **5b**, **5c**, **5d**, **15a**, **15b**, **15c**, and **15d** that transmit 250Mbps each.

Each transceiver **5a**, ..., **5d**, **15a**, ..., **15d** is connected to one end of one of four pair of unshielded twisted-pair cable **10a**, **10b**, **10c**, and **10d**. The transmitter **2** of each transceiver **5a**, ..., **5d**, **15a**, ..., **15d** forms a five level pulse amplitude modulated (PAM-5) shaped pulse signal that is transferred through the hybrid network **6** to one of the unshielded twisted-pair cable **10a**, ..., **10d**. The transmitted signal traverses the unshielded twisted-pair cable **10a**, ..., **10d** and is transferred through the hybrid network **6** to the receiver **4**. The received signal is sensed, retimed, equalized and transferred to other circuitry for extraction of the digital data.

The full-duplex bi-directional transmission consists of transmitting and receiving data simultaneously in both directions on each of the four wire pairs, minimizing the symbol rate (and thus, the occupied signal bandwidth) on each wire pair by one half, as compared to unidirectional transmission and reception. The hybrid network **6** is used to enable bi-directional transmission over single wire pairs by filtering out the transmit signal at the receiver. The hybrid network

6 has good trans-hybrid loss to minimize the amount of transmitter signal that is coupled into the receiver **4**, but it still cannot remove all of the transmitted signal from the adjacent transmitter **2**. The residual transmitted signal from the adjacent transmitter **2** from the hybrid **6** is defined as the transmit echo signal **25**.

Since the unshielded twisted-pair cable **10a**, **10b**, **10c**, **10d** are placed in close proximity to each other, often within the same cable, the crosstalk **20a**, **20b**, and **20c** from transmitters within the same gigabit Ethernet node are coupled to the receiver 2. The near-end crosstalk **20a**, **20b**, and **20c** and the transmit echo signal **25** must be cancelled from the receive signal to permit recovery of the transmitted signal.

Refer now to Fig. 2 for a more detailed discussion of the structure of a gigabit Ethernet node at one end of a bundled cable **10** containing four unshielded twisted-pair cable **10a**, ..., **10d**. Each transmitter **4a**, **4b**, **4c**, and **4d** receives an encoded and scrambled symbol to be transmitted from the side-stream scrambler and symbol encoder **30**, which in turn has received the digital data to be transferred from the Gigabit Media Independent Interface (GMII) **75**. The digital transmit filter **35** shapes the symbol to be transmitted to condition the transmitted spectrum. The digital-to-analog converter **40** creates the five level pulse amplitude modulation signal that is then transferred to the hybrid network **6a** for transmission on the unshielded twisted-pair cable **10a**.

A similar pulse amplitude modulated signal is simultaneously transmitted from a transmitter connected at the opposite end of the unshielded twisted-pair cable **10a**. The received signal is separated from the transmitted signal in the hybrid network **6a** and is the input to the analog-to-digital converter **45**. The digitized received signal is the feed-forward equalizer (FFE) **50** to compensate for signal distortion introduced in the communication channel. The feed-forward equalizer **50** combined with a feedback equalizer or decision feedback equalizer **80** often provides better signal equalization than linear equalization when the transmission medium (cable **10**) introduces strong signal attenuation with specific frequency regions. The feed-forward equalizer **50** does not modify the noise (echo, crosstalk, etc.).

The input signals from the side-stream scrambler and symbol encoder **30** to each transmitter **2b**, **2c**, and **2d** are the inputs to the near-end crosstalk cancellers **55a**, **55b**, and **55c** and to the echo canceller **60**. The near-end crosstalk cancellers **55a**, **55b**, and **55c** and the echo canceller **60** reproduce the near-end echo interference and the near-end crosstalk interference that is present in the received signal. The outputs of the near-end crosstalk cancellers **55a**, **55b**, and **55c** and the echo canceller **60** are the inputs to the summing circuit **65**. The equalized digitized received signal is transferred from the feed-forward equalizer **50** to the summing circuit **65**. The summing circuit **65** combines the reproduction of the echo interference and the near-end crosstalk interference and the equalized digitized received signal to cancel the echo

interference signal and the near-end crosstalk signals induced to the received signal as described above.

The Viterbi decoder and side-stream descrambler **70** provide error correction and resequencing of the receive signal to recover the digital data that is transferred to the Gigabit Media Independent Interface (GMII) **75** for further processing.

The near-end echo canceller **60** and the crosstalk cancellers **55a**, **55b**, and **55c** are known in the art and have been applied to applications such as **100** Base-T Ethernet, asynchronous transfer mode (ATM), local area networks (LAN), and telephone communication networks.

U. S. Patent 4,995,104 (Gitlin) describes a receiver that includes an interference canceller, which receives a corrupted signal and makes an estimate of the desired signal, subtracts the estimated desired signal from a delayed version of the received signal to form an estimate of the interference signal, then forms a final estimate of the desired signal by subtracting the estimated interference from a second delayed version of the received signal.

U. S. Patent 5,329,586 (Agazzi) teaches an echo canceling circuit and associated method for canceling errors encountered in data communications

decomposing a lookup-table nonlinear echo canceller into a plurality of smaller lookup tables, and combining outputs of the lookup tables.

U. S. Patent 5,887,032 (Cioffi) discusses a method and apparatus for crosstalk cancellation (e.g., NEXT interference) from received signals on a line by adaptively estimating the crosstalk interference from the other lines having interfering transmissions and by canceling the crosstalk interference using the estimated crosstalk interference.

U. S. Patent 4,669,116 (Agazzi) discloses an echo cancellation circuit for use with full-duplex data transmission systems. The echo canceller can operate in spite of time invariant non-linearities in the echo channel or in the implementation of the echo canceller itself (such as in D/A converters).

"A Pipelined Adaptive NEXT Canceller," Im, et al., IEEE Transactions on Signal Processing, pp. 2252 - 2258, Aug. 1998 Vol. 46 Issue: 8 ISSN: 1053-587X describes a near-end crosstalk (NEXT) canceller using a fine-grain pipelined architecture.

"100BASE-T2: 100 Mbit/S Ethernet Over Two Pairs Of Category-3 Cabling" Cherubini, et al., 1997 IEEE International Conference on Communications, pp. 1014 - 1018, 1997 Vol. 2, discusses the 100BASE-T2 physical layer specification for the receivers, particularly the adaptive digital

filters that are required for echo and NEXT cancellation, equalization, and interference suppression.

As described above, near-end echo comes from the imperfections of the near-end hybrid network **6a** that separates received signal from the transmitted signal and takes the significant part of the overall signal energy. The near-end echo signal degrades the receiver performance to a great extent and is a large source of error in the received signal. Therefore, the near-end echo canceller **60** is used to reduce the echo signal in the received signal. An adaptive echo cancellation technique is normally used because of its superior performance. For the receivers like gigabit Ethernet receivers, the performance requirement for the echo canceller is severe, and robustness of the received signal processing is required. Near-end crosstalk (NEXT), as described above, comes from the cross-coupling of the unshielded twisted-pair cable **10a**, **10b**, **10c**, and **10d** within a cable bundle and is one of the major sources of noise for the received signal.

In order to insure that the near-end echo and near-end crosstalk are minimized, a separate near-end echo/near-end crosstalk cancellation technique is necessary. The cancellation technique must not interact with other circuits in the receiver. In a receiver adopting a randomizing scrambler in the side-stream scrambler and symbol encoder **30**, a correlator maybe used for the echo cancellation as shown in Fig. 3. A correlator **100** obtains the echo signal

response for the transmitted symbols **X(k) 115**. Once the echo/NEXT signal response is acquired, a Finite Impulse Response (FIR) filter **105** set with the coefficients $C_0, \ldots, C_j$ **125** obtained by the correlator **100** generates the duplicated echo **e (k) 135** and subtracts the duplicated echo/NEXT signal **e(k) 135** from the received signal **X(k) 115** at the receiver. The coefficients $C_0, \ldots, C_j$ **125** at the FIR **105** may be updated once they are set. However, the coefficients $C_0, \ldots, C_j$ **125** at the FIR **105** are not usually updated once set because it usually takes thousands of symbols to generate new coefficients $C_0, \ldots, C_j$ **125** from the received signal **X(k) 115** in the correlator.

In a time varying echo/NEXT channel, such as the gigabit Ethernet, fixed coefficients $C_0, \ldots, C_j$ **125** cannot serve the purpose of the echo/NEXT cancellation since they cannot reflect the changes of the channel characteristics. Therefore, an update method for the coefficients $C_0, \ldots, C_j$ **125** is necessary. A windowed measurement for the coefficients $C_0, \ldots, C_j$ **125** is possible, but a more preferred way is to update the coefficients $C_0, \ldots, C_j$ **125** to reflect the most recent parameters of the channel.

Refer now to Fig. 4 for a discussion of the correlator **100**. Each coefficient $C_0, \ldots, C_j$ **125** of the FIR filter **105** is the normalized product of the received signal and a delayed version of the transmitted symbol **b(k)** ~~**20**~~ 120. That is

$$C_J(k) = N\big(x(k) * b(k - j)\big)$$

where:

C_j is the value of each coefficient **125** of the FIR filter **105**.

N is a normalizing factor for the coefficients **125**.

X(k) is the magnitude of the received signal **115**.

b(k-j) is the magnitude of the delayed transmitted signal **120**.

The multiplier circuit **150a**, **150b**, **150c**, ..., **150d** receives the received signal **X(k) 115** and the delayed transmitted signals **b(k-j) 165a**, **165b**, **165c**, ..., **165d**. The delayed transmitted signal **b(k-j) 165a**, **165b**, **165c**, ..., **165d** is the transmitted signal **b(k) 120** successively delayed through each of the unit delay elements **150a**, **150b**, **150c**, ..., **150d**. The product of the received signal **X(k) 115** and the delayed transmitted signal **165a**, **165b**, **165c**, ..., **165d** is the output of the multiplier **150a**, **150b**, **150c**, ..., **150d** and the input of the normalization circuit **160a**, **160b**, **160c**, ..., **160d**. The product is normalized to form the coefficients C_0, C_1, C_2, ..., C_j **125**, by summing a large number of products and dividing the sum of the squares of the delayed transmitted signal **b(k-j)**. Thus, each of the coefficients is determined as:

$$C_J = \frac{\sum_{k=0}^{n} (x(k) * b(k-j))}{\sum_{k=0}^{n} (b(k-j))^2}$$

where;

n is the number of symbols used to determine the coefficient C_j.

In the conventional prior art, the number of symbols **n** used to determine the coefficient C_j is 1024. Therefore, for example, the first coefficient C_0 becomes

$$C_0 = \frac{\sum_{k=0}^{1023} (x(k) * b(k))}{\sum_{k=0}^{1023} (b(k))^2}$$

Summary of the Invention

An object of this invention is to provide a communication transmitting and receiving system in which the effects of near-end echo and near-end crosstalk from the communication medium are mitigated.

Another object of this invention is to provide an apparatus, which is included in a communication receiving system for adaptively reproducing the near-end echo and near-end crosstalk signal, which are then subtracted from the received signal to mitigate the effects of the near-end echo and near-end crosstalk.

Further, another object of this invention is to adaptively generate filter coefficients for a Finite Impulse Response filter to reproduce the near-end echo and near-end crosstalk.

To accomplish the above objects, an apparatus to cancel echo and crosstalk interference in a receiver of a communication system having simultaneous transmission and receiving on a communication medium includes an adaptive correlator and a finite impulse response filter. The adaptive correlator is connected to a receiving circuit to acquire received signals from the communication medium. Further, the adaptive correlator is connected to at least one of a plurality of transmission channels of the communication system to acquire at least one transmitted signal from the adjacent transmission channels. The adaptive correlator generates a reproduction of the echo response of the received signals to near-end echo and near-end crosstalk interference from the transmitted signals at an arrival of each received signal.

The finite impulse filter is connected to the receiving circuit to acquire the received signals, and is connected to the adaptive correlator to receive a plurality of filter coefficients. The finite impulse filter reproduces the near-end echo and near-end crosstalk signals from the received signals based on the values of the plurality of filter coefficients. The reproduced near-end echo or near-end

crosstalk signals are combined with the received signals to cancel any echo and crosstalk interference from the received signals;

The filter coefficients are regenerated in the adaptive correlator at the arrival of each received signal and whereby each new filter coefficient is a weighted sum of a previous coefficient and one received signal multiplied by a time delayed version of one transmitted signal.

A first embodiment of the adaptive correlator has at least one first delaying means connected to one of the transmission channels to delay one of the transmitted signals. At least one first multiplying means is connected to the receiving circuit and one of the first delaying means to multiply the received signal by one delayed transmitted signal to produce a product of the received signals and the one delayed transmitted signal. At least one second multiplying means is connected to one of the first multiplying means to receive the product of the received signals multiplied by the one delayed transmitted signal and multiply this product by a first weighting factor ($\frac{\beta}{\sigma^2}$) to produce a first intermediate coefficient factor.

The adaptive correlator has at least one second delaying means to delay and retain the previous coefficient. At least one third multiplying means is connected to one second delaying means to multiply the previous coefficient by one minus a second weighting factor (β) to produce a second intermediate

coefficient factor. One of the second multiplying means and one of the third multiplying means is connected to one of a group of at least one summing means to add the first intermediate coefficient factor and the second intermediate coefficient factor to produce one new filter coefficient. The first weighting factor is a quotient of the second weighting factor divided by a variance (σ^2) of the transmitted signal. The second weighting factor β is chosen in a manner similar to an equivalent weighting factor used in to what is termed a leaky recursive least squares method to calculate the coefficients of an adaptive filter. For this embodiment of this invention, a number of the symbols **n** (for instance, **n**=256 symbols) is chosen and the weighting factor β is equal to the inverse of the number of the symbols. That is:

$$\beta = \frac{\mathbf{1}}{\mathbf{n}} = \frac{\mathbf{1}}{\mathbf{256}} \cong \mathbf{.004}$$

The received signals and the transmitted signals are digitized to form binary numbers indicating magnitudes of samples of the received signal and the transmitted signal.

A second embodiment of the adaptive correlator has at least one first delaying means connected to one of the transmission channels to receive and delay on of the digitized samples of the transmitted signal and at least one first shifting means connected to the receiving circuit and one of the first delaying means to shift one of the digitized samples of the received signal according to the binary number of the one digitized sample of the delayed transmitted signal

to produce the product of one received signal and the one time delayed transmitted signal. The adaptive correlator has at least one second shifting means to shift the product of the one received signal and the one time delayed transmitted signal by a first weighting factor ($\frac{1}{\sigma^2}$) to form a weighted product of the one received signal and the one time delayed transmitted signal.

Further, the adaptive correlator has at least one second delaying means to delay and retain the previous coefficient. One adder/subtractor means of a group of at least one adder/subtractor means is connected to one second shifting means and one second delaying means to combine the previous coefficient with the weighted product to form a partially weighted sum. One third shifting means of a group of at least one third shifting means is connected to one adder/subtractor to shift the partially weighted sum by a second weighting factor (β) to form an intermediate weighted sum. One adding means of a group of at least one adding means is connected to one second delaying means and to one third shifting means to additively combine the partially weighted sum and the intermediate weighted sum to generate the new filter coefficient.

In the second embodiment, the first weighting factor is the inverse of the variance (σ^2) of the transmitted signal and the second weighting factor β, as described above, is chosen in a manner similar to an equivalent weighting factor used in to what is termed a leaky recursive least squares method to calculate the coefficients of an adaptive filter. For this embodiment of this invention, a number

of the symbols **n** (for instance, **n**=256 symbols) is chosen and the weighting factor β is equal to the inverse of the number of the symbols. That is:

$$\beta = \frac{1}{n} = \frac{1}{256} \cong .004$$

Brief Description of the Drawings

Fig. 1 is a simplified diagram of the basic structure of a gigabit Ethernet communication link of the prior art.

Fig. 2 is a block diagram of a transceiver of a gigabit Ethernet communication system of the prior art.

Fig. 3 is a block diagram of a near-end echo and near-end crosstalk canceller of the prior art.

Fig. 4 is a schematic diagram of a correlator of the near-end echo and near-end crosstalk canceller of the prior art.

Fig. 5 is a schematic diagram of a first embodiment of a correlator of a near-end echo canceller and a near-end crosstalk canceller of this invention.

Fig. 6 is a schematic diagram of a second embodiment of a correlator of a near-end echo canceller and a near-end crosstalk canceller of this invention.

Figs. 7a-7c are flow diagrams of embodiments of the method to cancel near-end echo/near-end crosstalk from a received signal in a communication system of this invention

Detailed Description of the Invention

To meet the severe receiver requirements of the gigabit Ethernet, the near-end echo/near-end crosstalk canceller of this invention has a correlator **100** of Fig. 3 that creates each coefficient $C_0, \ldots, C_j$ **125** of the FIR filter **105** that is the function of the previous coefficient. Each coefficient $C_0, \ldots, C_j$ **125** is the weighted sum of the previous coefficient and the received signal **X(k) 115** multiplied by a time delayed version of the transmitted symbol **b(k) 120** and is summarized as follows:

$$C_J(k+1) = (1-\beta) * C_J(k) + \frac{\beta}{\sigma^2} * x(k) * b(k-j) \qquad \textbf{EQ. 1}$$

where:

$C_j(k)$ is the previous coefficient.

$Cj(k+1)$ is the filter coefficient for the FIR filter for the next received signal.

X(k) is the present received signal.

b(k-j) is the transmitted signal delayed by **j** delay units.

σ^2 is the variance of the transmitted symbol. For gigabit Ethernet $\sigma^2 = 2$.

β is a weighting factor.

The weighting factor β is chosen in a manner similar to an equivalent weighting factor used in to what is termed a leaky recursive least squares method to calculate the coefficients of an adaptive filter. For this embodiment of this invention, a number of the symbols **n** (for instance, **n**=256 symbols) is chosen and the weighting factor β is equal to the inverse of the number of the symbols. That is:

$$\beta = \frac{\mathbf{1}}{\mathbf{n}} = \frac{\mathbf{1}}{\mathbf{256}} \cong \mathbf{.004}$$

Refer now to Fig. 5 for a discussion of a first embodiment of the correlator **100** circuit of the near-end echo/near-end crosstalk cancellation circuit of this invention. The received signal **X(k) 115** and the delayed transmitted signals **b(k-j) 240a**, **240b**, **240c**, **...**, **240d** are the inputs to each of the first multipliers **200a**, **200b**, **200c**, **...**, **200d**. The delayed transmitted signals **b(k-j)** are the outputs of each of the unit delay elements **205a**, **205b**, **205c**, **...**, **205d**. The unit delay elements **205a**, **205b**, **205c**, **...**, **205d** successively delay the transmitted signal **b(k) 120** to form the delayed transmitted signals **b(k-j) 240a**, **240b**, **240c**, **...**, **240d**.

The output products **245a**, **245b**, **245c**, ..., **245d** and the first weighting factor ($\frac{\beta}{\sigma^2}$) **215** are inputs to the second multipliers **210a**, **210b**, **210c**, ..., **210d**. The outputs of the second multipliers **210a**, **210b**, **210c**, ..., **210d** form the weighted products **250a**, **250b**, **250c**, ..., **250d**.

Each of the new filter coefficients **C_0 (k +1)**, ..., **Cj(k+1) 225a**, **225b**, **225c**, ..., **225d** is the input to each of the unit delay elements **230a**, **230b**, **230c**, ..., **230d** to form the previous filter coefficients **255a**, **255b**, **255c**, ..., **255d**. A second weighting coefficient (β) **240** and the previous filter coefficients **255a**, **255b**, **255c**, ..., **255d** are the inputs to the third multipliers **235a**, **235b**, **235c**, ..., **235d**. The outputs of the third multipliers **235a**, **235b**, **235c**, ..., **235d** form the weighted previous filter coefficients **260a**, **260b**, **260c**, ..., **260d**.

The weighted products **250a**, **250b**, **250c**, ..., **250d** and the weighted previous filter coefficients **260a**, **260b**, **260c**, ..., **260d** are respectively the inputs to the summing circuits **220a**, **220b**, **220c**, ..., **220d** to be additively combined to form the new filter coefficients **Cj(k+1) 225a**, **225b**, **225c**, ..., **225d**. The new filter coefficients **Cj(k+1) 225a**, **225b**, **225c**, ..., **225d** are placed at the inputs **125** of the FIR filter **105** of Fig. 3 to set the FIR filter **105** to reproduce the near-end echo/near-end crosstalk for the received signal **X(k) 115** at a next instant.

It can be shown that the function of EQ. 1 can be rewritten to the form:

$$C_J(k+1) = C_j(k) + \beta\left(\frac{x(k) * b(k-j)}{\sigma^2} - C_j(k)\right) \qquad \textbf{EQ. 2}$$

As described above, the weighting factor β is chosen in a manner similar to an equivalent weighting factor used in to what is termed a leaky recursive least squares method to calculate the coefficients of an adaptive filter. For this embodiment of this invention, a number of the symbols **n** (for instance, **n**=256 symbols) is chosen and the weighting factor β is equal to the inverse of the number of the symbols. That is:

$$\beta = \frac{1}{n} = \frac{1}{256} \cong .004$$

Since the transmit signals **b(k) 120** and the received signals **X(k) 115** are digitized samples of the signals transmitted and received on the communication medium (cable **10** of Fig. 2), the multiplication can be performed with shift registers. Refer now to Fig. 6 for a discussion of a correlator **100** of a second embodiment of the near-end echo/near-end crosstalk canceller of this invention.

The digital form of the transmitted symbol **b(k)** is the input to each of the first unit delay elements **305a**, **305b**, **305c**, ..., **305d** to form the delayed transmitted symbols **340a**, **340b**, **340c**, ..., **340d**. The sampled digitized received signal **115** and the delayed transmitted signals **340a**, **340b**, **340c**, ..., **340d** are the inputs to the first shifters **300a**, **300b**, **300c**, ..., **300d**. The first shifters **300a**, **300b**, **300c**, ..., **300d** shift the sampled digitized received signal **115** according to

the values of the delayed transmitted signals **340a**, **340b**, **340c**, ..., **340d** to form the output products **345a**, **345b**, **345c**, ..., **345d**.

The output products **345a**, **345b**, **345c**, ..., **345d** and a first weighting factor **315** are the inputs respectively to the second shifters **310a**, **310b**, **310c**, ..., **310d**. The first weighting factor **315** in this embodiment is the inverse of the variance (σ^2) of transmitted symbols. As described before, for gigabit Ethernet the variance of the transmitted symbols (σ^2) is 2.

The second shifters **310a**, **310b**, **310c**, ..., **310d** each shift the output products **345a**, **345b**, **345c**, ..., **345d** according to the binary value of the first weighting factor ($\frac{1}{\sigma^2}$) to form the weighted products **350a**, **350b**, **350c**, ..., **350d**.

The new filter coefficients **Cj(k+1) 325a**, **325b**, **325c**, ..., **325d** are the inputs to the second delay elements **330a**, **330b**, **330c**, ..., **330d**. The new filter coefficients **Cj(k+1) 325a**, **325b**, **325c**, ..., **325d** are delayed and retained for one timing cycle to become the previous filter coefficients **355a**, **355b**, **355c**, ..., **355d**.

The weighted products **350a**, **350b**, **350c**, ..., **350d** and the previous filter coefficients **355a**, **355b**, **355c**, ..., **355d** are subtractively combined in the arithmetic combining circuits **370a**, **370b**, **370c**, ..., **370d** to form the preliminary weighted sums **375a**, **375b**, **375c**, ..., **375d**. The preliminary weighted sums

375a, **375b**, **375c**, ..., **375d** and the previous coefficients **355a**, **355b**, **355c**, ..., **355d** are the input to the summing circuits **320a**, **320b**, **320c**, ..., **320d** where they are additively combined to form the new filter coefficients **C_0(k+1)**, **C_1(k+1)**, **C_2(k+1)**, ..., **Cj(k+1) 225a**, **225b**, **225c**, ..., **225d**. The new filter coefficients **C_0(k+1)**, **C_1(k+1)**, **C_2(k+1)**, ..., **Cj(k+1) 225a**, **225b**, **225c**, ..., **225d**, as stated prior, are the inputs **125** to the FIR filter **105** of Fig. 3, to set the FIR filter **105** to reproduce the near-end echo/near-end crosstalk for the received signal **X(k) 115** at the next digitized sample of the received signal **X(k) 115**.

Since the transmit symbols and the received signals are digital words that respectively are inputs to the digital-to-analog converter **40** of Fig. 2 and outputs from the analog-to-digital converter **45** of Fig. 2, it is possible to perform the reproduction of the near-end echo/near-end crosstalk as digital words within a computer system such as a digital signal processor. A method to cancel near-end echo interference and near-end crosstalk interference that is implemented in a digital signal processor is illustrated in Figs. 7 a-c.

A method to cancel echo interference and crosstalk interference present in a received signal from a communication medium begins with acquiring **400** the received signal with the echo interference and the crosstalk interference and acquiring **405** the transmitted signals that generate the echo interference and the crosstalk interference. The transmitted signal is then delayed **410** to form a delayed transmitted signal. The received signal is then correlated with at least

one transmitted signal at each cycle of the received signal to generate **415** a new filter coefficient as a weighted sum of a previous filter coefficient and the received signal multiplied by the delayed transmitted signal. The received signal is then filtered **420** to reproduce the echo interference and the crosstalk interference. The reproduction of the echo interference and the crosstalk interference are received combined **425** with the received signal to cancel the echo interference and the crosstalk interference in the received signal.

A first embodiment of the correlating of the received signal begins with at least one transmit signal to generate **415** the new filter coefficients begins by multiplying the delayed signal by at least one of the transmitted signals to form a first product. The first product is then multiplied **435** by a first weighting factor ($\frac{\beta}{\sigma^2}$) to form a first weighted product.

The previous filter coefficient are multiplied **440** by a second weighting factor (**1**-β) to form a second weighted product. The first weighted product and the second weighted product are then summed **445** to form the weighted sum. The first weighting factor is the quotient of the factor (β) divided by the variance of the transmitted signal (σ^2). As stated above the variance (σ^2) of the transmitted signal for gigabit Ethernet is 2. The second weighting factor is one minus the factor (β).

As previously described, the weighting factor β is chosen in a manner similar to an equivalent weighting factor used in to what is termed a leaky recursive least squares method to calculate the coefficients of an adaptive filter. For this embodiment of this invention, a number of the symbols **n** (for instance, **n**=256 symbols) is chosen and the weighting factor β is equal to the inverse of the number of the symbols. That is:

$$\beta = \frac{1}{n} = \frac{1}{256} \cong .004$$

A second embodiment of the correlating the received signal with at least one transmit signal to generate **415** the new filter coefficient begins by shifting **450** the received signal according to the delayed transmit signal to form a product of the received signal and the delayed transmit signal. The product of the received signal and the delayed transmit signal is further shifted **455** according to a first weighting factor ($\frac{1}{\sigma^2}$) to form a first weighted product. The first weighted product is additively combined with the previous filter coefficient to form a first preliminary weighted sum.

The first preliminary weighted sum is then shifted **470** by a second weighting factor (β) to form a second weighted product. The second weighted product and the previous filter coefficient are then additively combined **475** to generate the new filter coefficient.

In the second embodiment, the first weighting factor is the inverse of the variance (σ^2) of the transmitted signal and the second weighting factor is the factor β.

The second weighting factor β is chosen, as above described, in a manner similar to an equivalent weighting factor used in to what is termed a leaky recursive least squares method to calculate the coefficients of an adaptive filter. For this embodiment of this invention, a number of the symbols **n** (for instance, **n**=256 symbols) is chosen and the weighting factor β is equal to the inverse of the number of the symbols. That is:

$$\beta = \frac{\mathbf{1}}{\mathbf{n}} = \frac{\mathbf{1}}{\mathbf{256}} \cong \mathbf{.004}$$

While this invention has been particularly shown and described with reference to the preferred embodiments thereof, it will be understood by those skilled in the art that various changes in form and details may be made without departing from the spirit and scope of the invention.

The invention claimed is: